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EUROPEAN PATENT APPLICATION

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(54) Electromagnetic actuator system.

(57) Using as an actuator a stepper motor (11, 12) operated in a position-sensed closed loop mode via an electronic commutator by position and anomaly corrected motion control signals provided by a programmable logic array (14) and by a micro-processor (24), provides a precision power positioner. The correct enabling of the motor coils, based on position sensing of the motor phase, using electronic commutation for the stepper motor, transforms the problem of enabling coil selection to the programmable logic array, which by its logic provides a simplified operating control as a function of present armature (12) position feedback (21), and by a micro-processor corrects for anomalies such as temperature rise, square-law, and phase increase as a function of position desired and present armature position feedback signals (22, 23), and thus provides high positional accuracy, high speed and force (torque) control independent of the motor cardinal step positions. A standard 3-phase magnetic resistor sensor (13) is used both to sense the motor phase and to feed back the motor position to the micro-processor for calculation of corrections; actual control of the stepper motor is communicated as a single amplitude modulated control signal.

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FIG. 6

STEPPER MOTOR CONTROLLER
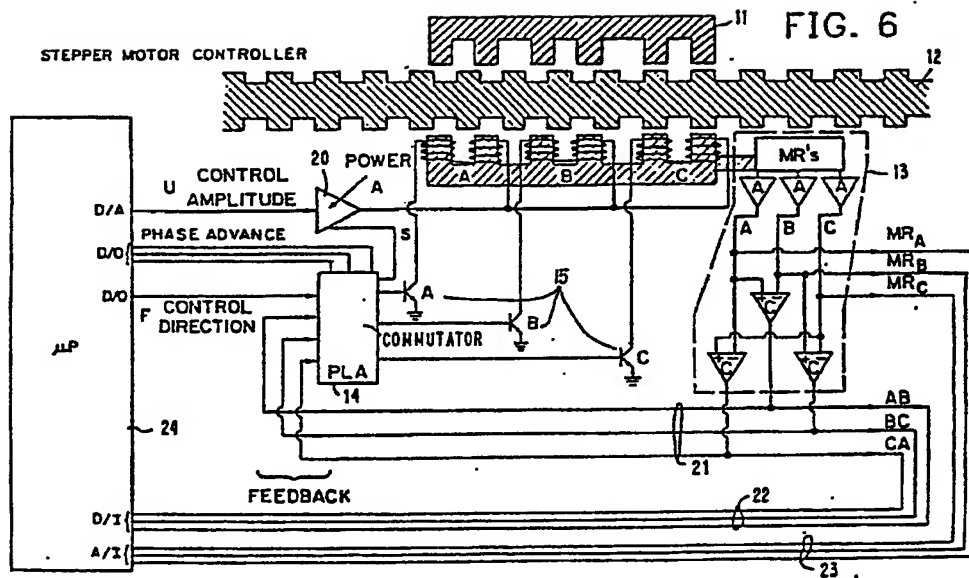
Electromagnetic Actuator System

This inventi n relates to an electromagnetic actuator system, and more particularly relates to an electromagnetic actuator system using a stepper motor.

5 To generate torque an ordinary D.C. motor has many windings which are automatically switched on or commutated at the appropriate time by brushes, or in a brushless D.C. motor by a position sensor and electronic switches.

10 Stepping motors have a high force or torque output. Since they can produce this torque at very low speeds they are quite attractive as direct drive actuators for robots. Unfortunately, the common "control" schemes for stepping motors use sequencing logic which can only position the motors at their cardinal step positions. This conven-
15 tional control does not operate well at high speeds or with varying loads.

For a direct drive application it is desirable that the stepping motor position itself between cardinal steps. The usual approach of "micro-
20 stepping", providing equilibrium point between cardinal step positions through a balancing choice of coil currents, is cumbersome, doesn't account for motor dynamics, and is wasteful of power. Secondly, control of delivered torque or force is a desirable trait in a r bot actuator, but such control is difficult with conventional step er
25 motor control laws, such as Proportional Derivative (PD) and Proportional Integral Derivative (PID).

The following are representative of the prior art:

30 Publications

L. Thomson and M. Lee, "Universal Brush-less Motor Commutator", Proc of Thirteenth Annual Symp on Incremental Motion Control Systems and Devices, University of Illinois, May 1984.

35

C. Oudet, "DC Versus Stepper: A New Approach", Proc of Thirteenth
Annual Symp on Incremental M tion Control Systems and Devices,
University of Illinois, May 1984, shows how to convert a permanent
magnet stepper motor to a DC motor, but still the implementation

5 requires motor motion to detect the motor phase sensor and the coil
switching circuits. There is a discussion of the basic theory of
permanent magnet stepper motors.

R. Welburn, "Ultra High Torque System for Direct Drive Robotics",
10 Robot 8 Conference, Detroit USA, June/1984, shows a synchronous motor
with electronic commutator and twelve bit digital position feedback,
but does not have square-root compensation and uses a complex com-
mutator.

15 W.R. Pearson and Paresh C. Sen, "Brushless DC Motor Propulsion Using
Synchronous Motors for Transit Systems", IEEE Transactions on Indus-
trial Electronics, Vol. IE-31, No. 4, November 1984, shows an elec-
tronic commutator for synchronous motors.

20 J.P. Pawletko, "Dynamic Responses and Control Aspects of Linear
Stepping Motors", Proc of Fifth Annual Symp on Incremental Motion
Control Systems and Devices, University of Illinois, May 1976, shows
an experimental motor and summarizes control techniques in stepper
motors.

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J.P. Pawletko and H.D. Chai, "Linear Stepping Motor with Uncoupled
Phases", Proc of Thirteenth Annual Symp on Incremental Motion Control
Systems and Devices, University of Illinois, May 1984, shows another
experimental motor.

30

The RRA card (Robotic Research Attachment) is a Motorola 68000 micr-
processor and peripheral board designed to attach to the IBM Series .

B.C. Kuo, "Theory and Applications of Step Motors", West Publishing
35 Co., 1974.

R.H. Taylor, P.D. Summers and J.M. Meyer, "AML: A Manufacturing Language", International Journal of Robotics Research, Vol. 1, No. 3, Fall 1982.

5 Patents

3,548,274, Mako, STEPPING MOTOR SPEED CONTROL, December 15, 1970, shows a capacitive electronic commutator for a stepping motor.

10 3,636,429, Jakubowski et al., LOGIC CIRCUITRY FOR PROVIDING STEPPING CONTROL FOR STEPPING MOTORS, January 18, 1972, shows a closed loop system for controlling the stopping sequence in stepper motor operation.

15 3,766,459, McIntosh et al., ELECTRONIC STEPPING DRIVE FOR DC TORQUE MOTOR, October 16, 1973, shows operation of a DC servomotor using a digital to analog converter to convert step pulses to analog amplitude input.

20 4,025,837, Meier et al., ADAPTIVE CONTROL CIRCUIT FOR A STEPPING MOTOR, May 24, 1977, shows a technique to move in minimum time by controlling the proportion of acceleration signals to deceleration signals, using position feedback summed with the analog input to avoid overshoot.

25

4,042,868, Rhodes, STEPPER MOTOR CONTROL APPARATUS, uses position feedback in a motor control circuit with duty cycle control.

Reissue 31,229 (4,136,308) King, STEPPING MOTOR CONTROL, May 3, 1983,
30 shows operation of a stepping motor using an electronic cummutatc ⁚ from a DC power source, using back EMF in the stator coils to sens ⁚ rotor position.

4,158,800, Jahelka et al., CONTROL SYSTEM, June 19, 1979, shows a
35 stepper motor with a closed loop feedback system, using a coded disk for position feedback, and providing for speed change by varying lead angle.

4,187,455, Martin, STEPPER MOTOR FEEDBACK IN POSITION SERVO LOOP, Feb. 5, 1980, shows a digitalized position feedback for subtraction from a digital position on command word so as to generate a position error signal to drive the stepper motor in a servo loop.

5

4,254,370, Miller, CLOSED LOOP POSITIONER FOR A STEPPING MOTOR DRIVEN BY A BUFFERED TRANSLATOR, March 3, 1981, shows the use of a buffered translator as a commutator, with position feedback, to operate the stepper motor at its inherent speed even though demand is for greater

10   speed.

4,349,771, Buxton, STEPPER MOTOR CONTROL CIRCUIT, September 14, 1982, shows a closed loop system using an integrator system to compare a demand signal with a shaft position signal.

15

4,426,608, Larson et al., STEPPING MOTOR CONTROL SYSTEM, January 17, 1984, shows modulation of lead angle in accordance with a selected duty cycle, using optical position sensing.

20   4,429,268, Yajima et al., SPEED CONTROL FOR STEP MOTORS, January 31, 1984, shows position feedback and electronic commutation in a closed loop speed control, using acceleration and deceleration control memory.

25   The prior art, while it shows various electronic commutator or direct electronic control schemes for position feedback closed loop control of motors, does not teach nor suggest the invention.

The present invention is disclosed in the attached claims.

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The object of the invention is to utilize the power and economy of stepping motor actuators in high precision assignments, with dynamic control by a single current (voltage) control amplitude.

35   Another object of the invention is to make possible a practical electronic commutator system for a stepper motor, with motor control

by a single control variable, the effect of which is position inde-
pendent, so that from a control view a stepper motor equipped with an
electronic commutator system according to the invention looks like a
standard D.C. motor.

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Another object of the invention is to simplify stepper motor control
to achieve control by a single control variable which is position
independent, while still permitting anomaly correction via the single
control variable.

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Another object of the invention is to apply electronic commutation to
stepping motors; to enable appropriate coils according to control law,
through a single control amplitude demand input.

15   An object of the invention is to make stepper motor control achievable
with a single control input, independent of motor position (angle),
and to obtain maximum force (torque) from the motor.

An object of the invention is to follow the commutation principle:

20

    Given the present position and speed of the motor and a desired
    direction of motor's force, energize the coils which maximize
    the force in the desired direction. For many motors this rule
    achieves a nearly position independent force.

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Another object of the invention is to achieve position independence,
by energizing the enabled coils with an appropriate magnitude.

Another object of the invention is to make the commutation principle
30   effective for all types of linear or rotary stepper motors, includ-ng
linear variable-reluctance stepper motors, coupled phase or uncoup-ed
phase, or permanent magnet stepper motors, or hybrids, using the
magnitude of current or voltage applied to the energized coils com-
bined with the desired direction of the motor's force to form the
35   control input, and using a control law to dictate the control input's
value.

A feature of the invention is the use f a programmable logic array in an electronic commutator to provide coil enabling selection signals as a composite function of present position feedback signals, in a first feedback loop, control direction signals, signals and phase advance
5 signals, using a micro-processor to calculate dynamic control amplitude signals and phase advance signals as a composite function of present position feedback signals, in a second or subsequent feedback loop, providing position signal control information, with dynamic position and anomaly correction forwarded as a current (voltage)
10 control amplitude signal to the motor coils of the stepper motor, and as phase advance signals to the programmable logic array.

An advantage of the invention is that an available stepper motor, known for power and economy, may be operated as a precision positioner
15 by the combination of position sensing and external electronics including a micro-processor, programmed logic array and electronic commutator.

Another advantage of the invention is that a stepper motor equipped
20 according to the invention may be positioned in increments of fractional steps, and held in place indefinitely with very small use of power.

Another advantage of the invention is that a stepper motor may be
25 operated according to complex dynamic control patterns corrected for present position, desired position, temperature and magnetic anomalies and other factors considered by a micro-processor, which control patterns may be easily reconfigured by programming.

30 The foregoing and other objects, features and advantages of the invention will be apparent from the more particular description of the preferred embodiments of the invention, as illustrated in the accompanying drawings.

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Brief Description of the Drawings

FIG. 1 is a composite position diagram of a stepper motor showing half-step control.

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FIG. 2 is a diagram showing uncoupled phase 3-phase linear variable reluctance (V.R.) motor laminations, showing two E-cores and a segment of the bar.

10 FIG. 3 is a diagram showing commutator parts for a 3-phase uncoupled phase stepper motor.

FIG. 4 is a phase diagram showing the principle of commutation of a 3-phase stepper motor.

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FIG. 5 is a phase diagram showing static force ripple of a 3-phase stepper motor.

FIG. 6 is a schematic diagram of a micro-processor controller of an
20 electronically commutated stepper motor according to the invention.

FIG. 7 is a position/time diagram of small step response using a PD controller.

25 FIG. 8 is a position/time diagram of large step response using a PD controller with two sets of gains.

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Description of a Preferred Embodiment of the Invention

The preferred embodiment is an electromagnetic actuator system using a double closed-loop dynamic anomaly corrected computer selected electronic commutation technique for control of a linear or a rotary

5 stepper motor, for the correct enabling of the motor coils based on position sensing of the motor phase, application of control law and anomaly correction, and selection of destination.

The electronic commutator for the stepper motor has the same function

10 as brushes in a D.C. motor and is analogous to the commutator of a brushless D.C. motor. The computer provides motion-to-destination profiles, corrected for present position and for anomalies such as a square-law in variable reluctance stepper motors, as control amplitude signals to the motor coils selected by the electronic commutator.

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In the electromagnetic actuator system of the invention, the stepper motor has the ability to function much like a D.C. servomotor. Commutation is the correct enabling of the motor coils determined by position sensing. Electronic commutation in the electromagnetic ac-

20 tuator system of the invention has the same function as brushes in a D.C. motor is analogous to electronic commutation of a brushless D.C. motor. Using electronic commutation and a simple control law (e.g. PID control), one can achieve high positional accuracy, high speed, and force or torque control of linear or rotary stepping motors indepen-

25 dent of the motor's cardinal step positions. These capabilities allow the use of stepper motors as efficient direct drive actuators for a robot.

This motor is composed of two parts, one moving relative to the other:

30

Coils   A motor includes two rigidly connected E-cores. On each E-core three coils are mounted, one on every leg.

Bar   The bar, situated between the two E-cores, has rectangular

35   teeth of a fixed period or pitch on its surface.

Figure 1 illustrates a 3-phase V.R. linear stepper motor. The motor
includes tow rigidly connected E-cores 11 each having three legs on
which coils are mounted. A bar 12 composed of rectangular teeth of a
fixed period or pitch is situated between the two opposing E-cores.

5   This particular bar has a tooth pitch of 2.54 mm (0.1 inch). Figure 1
    shows the result of sequencing the motor to move the bar to the right
    by half-steps. Six half steps are required to move the bar one tooth
    pitch. In half-stepping the bar to the right the coils are energized
    as shown in the figure by alternatively energizing one and two coils.

10 Motion to the left is achieved by reversing the sequence of coils
    energized. It is also possible to advance the motor by full-steps
    rather than half-steps by energizing only one coil at a time, but the
    greater variation in force makes this method less attractive. The six
    coarse positions of the motor in Figure 1 are referred to as the

15 commutation phase. The commutation phases and the desired direction of
    motion determine which coils are to be energized.

Figure 1 also illustrated the complicated flux paths of this coupled
phase motor. For example the magnetic flux from a single coil being

20 energized passes through all three legs of the E-cores.

Sequencing through phases 1, 2, 3, 4, 5 to 0 and back to 1 would
require flux direction in coil AA' to change from up to down. Since
such a change cannot happen instantaneously, it is better not to

25 require the flux's orientation to reverse. By alternately stepping the
    motor through the 6 phases with the flux oriented as in Figure 1 and
    then with the fluxes reversed, such abrupt changes in flux orientation
    are removed. The sequencing of this motor and many rotary motors is
    thus very complicated.

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    Figures 2 and 3 show an uncoupled motor, with simplified sequencing
    developed by Pawletko and Chai. The basic principles of operation of
    variable reluctance motors is the same in the uncoupled motor, but the
    relative direction of the flux in each of the motor's coils is un-

35 important. In the following discussion it is simpler first to discuss
    the commutation of the uncoupled motor and then to show how a coupled
    motor can also be commutated. Stepper motor 11-12 has position sensor
    13, which feeds back position information to Programmable Logic Array

(PLA) 14, which provides control signals to commutation switches 15. Switches 15 complete circuits for enabling currents to respectively connected coils 16, to accomplish stepper motor operation. The position sensor 13 is made up of magnetic resistors (MRs) 12, arranged to
5 face the teeth of bar 12 in a known relationship for proximity sensing. The MRs 17 provide outputs via amplifiers 18 to comparators 19, which thus provide binary signals for feedback.

Power amplifier 20 receives control amplitude stepper motor demand
10 signals and provides power control accordingly to the motor coils.

Feedback connectors provide a feedback path 21 to return position data to PLA 14.

15 Commutation Phase Sensing

As in a brushless D.C. motor commutator, a stepper motor commutator includes three parts: a position sensor, commutator logic, and electronic coil switches. The position sensor must provide enough information to detect which phase, 0, 1,...., or 5, the motor is in.
20

Figure 4 illustrates how three magnetic resistor (MR) sensors can efficiently provide the necessary signals. The MRs are mounted on an E-core and sense the changing permeance as the bar moves under them. Figure 4 shows three MR output voltages each of a period of one tooth
25 pitch, 2.54 mm (0.1 inch). Because each MR is physically mounted one third of a tooth pitch from its neighbor, the signals are 120 degrees apart in phase. Comparators, (C) in Figure 3, on the outputs of the MRs produce three binary signals, AB, BC, and CA. The three digit binary word AB, BC, CA represents a Gray code of the commutation phase
30 in correspondence with one of the positions the motor might be in as in Figure 1. Completely independent from deriving the phase information to control commutation. The MR voltages can be measured by an A/D converter and processed to determine the motor's position.

35 Figure 4 also shows the choice of coils to be enabled as a function of the six discrete phases and the desired force direction, Forward or Reverse.

Figure 5 illustrates the force each coil and pair of coils can produce
as a function of position with fixed coil current and the result of
commutation. For an approximation of perfectly sinusoidal forces as a
function of position - the true forces are not so perfect - commuta-
5 tion provides a nearly constant torque. The variation is only 13%.

The commutation logic for the 3-phase uncoupled stepper motor is
represented by the following three logical equations,

10
$$A = F \overline{BC} + \overline{F} BC$$

$$B = F \overline{CA} + \overline{F} CA \qquad (1)$$

$$C = F \overline{AB} + \overline{F} AB$$

15 where A, B, and C are logical variables that represent respectively if
coil A or B or C is on. AB, BC, and CA represent logical values of the
comparison between the outputs of the corresponding MRs and so provide
the commutation phase in a Gray code.

20 For a coupled motor the commutation logic is somewhat more complex, as
a result of the coupling between the motor's phases which requires the
motor to go through twelve half-steps to return to the same orienta-
tion of magnetic flux.

25 The sequence of coils enabled follows the typical open-loop sequence
of half-steps with an average phase lead of 1.5 phases (between 1 and
2 phases) in the direction of the motion. The main difference between
electronic commutation and usual half-stepping is that the coil
sequencing is generated automatically by closed-loop feedback from the
30 motor phase sensor (the MRs in our example). In fact the commutator
logic functions as a very fast closed-loop feedback system around the
motor (at 1.5 m/sec, 3 600 coil switches/sec at the right time are
necessary). In Figure 3 the coil switching circuits are represented by
"symbolic" transistors A, B, C. The actual circuits can be quite
35 complex depending on the maximum voltage used and switching times
required.

The practical limitation resulting from the switching circuits is a limit on the minimum time it takes the current through a coil to be switched on or off. This limit is imposed by the coil inductance, L, and the maximum voltage, V_{max}, the switching circuit can handle

5 through the equation $|di/dt|_{max} = V_{max}/L$.

The delay in the build up of current and force effectively causes the correct commutation to be delayed. This problem may be partially overcome by advancing the measured motor phase and thus advancing the

10 commutation. The results are the same as in exitation advance for conventional stepper motor control (7); the motor develops higher torque at high speeds.

The following is the required logic to achieve phase advance where V_s

15 is the sign (direction) of the required phase advance and V_1, V_0 is the binary magnitude in units of half-steps:

$$\tilde{AB} = V_s \, V_0 \, \overline{V}_1 \, \overline{BC} + V_s \, \overline{V}_0 \, V_1 \, CA +$$

20

$$V_0 \, V_1 \, \overline{AB} + \overline{V}_s \, V_0 \, \overline{V}_1 \, \overline{CA} +$$

$$\overline{V}_s \, \overline{V}_0 \, V_1 \, BC + \overline{V}_0 \, \overline{V}_1 \, AB$$

25

$$\tilde{BC} = V_s \, V_0 \, \overline{V}_1 \, \overline{CA} + V_s \, \overline{V}_0 \, V_1 \, AB +$$

30

$$V_0 \, V_1 \, \overline{BC} + \overline{V}_s \, V_0 \, \overline{V}_1 \, \overline{AB} +$$

$$\overline{V}_s \, \overline{V}_0 \, V_1 \, CA + \overline{V}_0 \, \overline{V}_1 \, BC$$

35

$$\overset{\sim}{CA} = V_s\ V_0\ \overline{V}_1\ \overline{AB} + V_s\ \overline{V}_0\ V_1\ BC +$$

$$V_0\ V_1\ \overline{CA} + \overline{V}_s\ V_0\ \overline{V}_1\ \overline{BC} +$$

$$\overline{V}_s\ \overline{V}_0\ V_1\ AB + \overline{V}_0\ \overline{V}_1\ CA \qquad\qquad (2)$$

The new binary word $\overset{\sim}{AB}$, $\overset{\sim}{BC}$, $\overset{\sim}{CA}$ represents the position phase measurement advanced in the direction of V_s an amount determined by the magnitude V_1, V_0. To arrive at the correct commutation including advance, $\overset{\sim}{AB}$, $\overset{\sim}{BC}$, $\overset{\sim}{CA}$ replaces AB, BC, CA in the commutator logic (1).

Control of a Commutated Step Motor

A digital controller for -an electronically commutated stepper motor need only define how much voltage (or current) is to be applied to the motor at an instant and the direction in which the force is to be generated by the motor. In an ordinary D.C. servo-motor system the same signal is provided by the magnitude and sign of the driving voltage to the D.C. motor.

Both coupled and uncoupled phase variable reluctance stepper motors have been driven with a digital PID (proportional, integral, derivative) control law. While this represents a simple servo control strategy, its use in controlling a stepping motors is new.

A Motorola 68000 with the digital and analog ports found on an RA card (8) was a convenient host for experiments, although numerous commercially available micro-processor systems would do. The processor is relegated the task of interpreting motor position, estimating motor velocity, and synthesizing the control signal (Figure 6).

Figure 6 is similar to Figure 3 but with the addition of coarse position feedback loop 22 and fine position feedback loop 23, which provide position data micro-processor 24.

Motor position can be derived from the MR signals or from a distinct encoder. Velocity could be measured from a tachometer, but the added expense and complexity, especially in the linear actuator case, was not justified for our application. The processor could provide an

5 adequate estimate of velocity.

A Simple PID Controller

The first step in computing the control term is to form the difference between the desired position, X_d, and the actual position of the motor

10 at time k, X_k. Given the velocity of the motor at time k as v_k, a discrete PD control law is then

$$U_k = - K_e (X_k - X_d) - K_v v_k \tag{3}$$

15 where K, and K_v are gains on positional error and velocity, respectively, and U_k is the control.

As no tachometer is present, an estimate, $\hat{v}_k$, must be used for v_k. The simplest estimate is to use the first backward difference, a numerical

20 derivative, of the past postition measurements,

$$\hat{v}_k = X_k - X_{k-1} \tag{4}$$

As this is known to be a rather "noisy" estimator, other simple

25 schemes are employed. For example, by fitting 2'nd or 3'rd order polynomials to the data X_k, X_{k-1}, X_{k-2}, ... and then evaluating the derivative of the polynomial at time k, the error in such an estimate can be proven smaller for sufficiently high sample rates. These estimates did in fact improve the performance of the PID servo run at

30 various sample rates between 0.5 and 2 kHz.

With $U_k = - K_e (X_k - X_d) - K_v \hat{v}_k$, the sign of U_k determines the direction the motor is to accelerate and the magnitude determines the magnitude of voltage. Hence the sign of U_k is output to the commutator

35 for the signal F, the magnitude of U_k is output to the power amplifier to drive the motor, and the value of $\hat{v}_k$ is quantized and provided to the commutator for control of the phase advance (Figure 6). Lastly, to

accommodate a zer steady state error criterion, a linear control
designer adds an integral term – a sum in the discrete case – to the
control, forming the PID control:

5

$$U_k = - K_e(X_k - X_d) - K_v \hat{v}_k$$

$$- K_i \sum_{j=n}^{k} (X_j - X_d) \tag{5}$$

10

where $n < k$ is chosen to be some finite value. If this control law is
asymptotically stable, the value of position X_k as k tends to infinity
is X_d exactly.

15   This simple control provides enormous flexibility. The gains may be
adjusted to specify the transient response of the motor depending on
load and desired trajectory. The entire motor system is easily incor-
porated into robot applications where motions may be specified by a
robot programming language like AML (9).

20

Compensation for Non-Linearity

V.R. motors, unlike D.C. servo motors, have a pronounced non-linear-
ity. At current levels well below flux saturating values, the static
force developed is proportional to the square of the coil current (5,
25   10). As we are trying to "servo" the motor to a desired position with
a linear control law, we might expect this non-linearity to adversely
affect the performance. The obvious compensation for the non-linearity
is not to drive the motor with U_k but rather with

30

$$u_k = sign(U_K) \sqrt{|U_K|} \tag{6}$$

This compensation is correct when current is used as the control. In
fact, far higher stiffness – larger values of K_e – with good transient
response could be achieved with the square root than without it even
35   when voltage was applied as the control. For better performance u_k
replaces U_k as the contr l signal driving the motor – commutator
system.

Performance

The good performance of this control scheme coupled with the commu-
tator makes the approach very attractive. For large moves, a PD
controller has driven the uncoupled phase motor at speeds over
5 1.4 m/sec (55 inch/sec) on a 0.5 m (18 inch) bar, and yet stopped it
in a critically damped manner. In another application a PID controller
provides zero error on a 0.002 mm (0.00008 inch) resolution encoder
with a good transient response.

10 Figure 7 shows the transient response while commanding the coupled
phase linear motor to move in 0.025 mm steps (0.001 inch). The appar-
ent stiffness of this PD controller (the force required to displace
the motor from various commanded equilibrium positions) was measured
at 90 Nt/mm (450 lb/inch). This stiffness is comparable with that of
15 micro-stepping methods.

Figure 8 shows the result of commanding a 300 mm (6 inch) move under
direction of AML at a speed of 0.5 m/sec (20 inch/sec) and accelera-
tion/deceleration of 10 m/sec/sec (1 g). The error between the posi-
20 tion goal, X_d, and the actual position at each instant hardly exceeds
5 mm (0.2 inch). To achieve this performance the gains K_v and K_e
switched between two values depending on the position error and the
velocity. Figure 1 shows the results of sequencing the motor to move
the bar to the right by half-steps - it is convenient here to refer to
25 the coils as fixed and the bar in motion. It is necessary to go
through the six half-steps shown to move the bar one tooth pitch as
indicated by the motion of the horizontal arrow in Figure 1. Thus,
every half-step moves the bar one sixth of a tooth pitch (e.g. about
0.43 mm).
30
In half-stepping the bar to the right, alternating between one and two
coils energized, the sequence of phases (half-steps) 1, 2, 3, 4, 5 and
0 is repeated. Motion to the left is achieved by reversing the se-
quence of phases. It is also possible to advance the motor by full-
35 steps rather than half-steps by the sequence of phases 1, 3 and 5
repeated, with only one coil energized at a time. This method is

inferior because of larger steps, lower maximum force, and the larger
force variation during the stepping of the motor.

The best positional resolution is one half-step (e.g. about 0.43 mm)
5 which is achieved by using the half-step method. Using the full-step
method only a full-step resolution can be achieved.

If the motor is sequenced through phases 1, 2, 3, 4, 5 to 0 and back
to 1 then the flux direction in coil AA' changes from up to down. Such
10 a change in flux direction cannot be done in zero time (with a finite
voltage), and thus, it is more desirable not to require the change in
flux direction at this transition. This can be achieved by a "double
cycle", alternating between two cycles of half-steps where the first
is with flux directions shown in Figure 1 while the second is with all
15 fluxes in the opposite direction. This complicates the sequencing of
such a magnetically coupled stepper motor.

An uncoupled motor developed by Pawletko and Chai has simplified
sequencing. The basic principle of operation of a variable reluctance
20 uncoupled motor is the same as a coupled motor, but in an uncoupled
motor the relative direction of the flux in each of the motor coils is
unimportant. Thus the "double cycle" sequencing needed in a coupled
motor is not required in an uncoupled motor.

25 The simplicity of sequencing an uncoupled motor is also reflected in a
particularly simple electronic commutator for an uncoupled motor. This
is expected for when the motor is moving in one direction the commuta-
tor reproduces (by closed-loop-feedback) the standard open-loop-half-
step motor sequencing. How such a closed-loop commutator works is
30 explained in the next Section.

Electronic Commutator Operation
Similar to a brushless D.C. motor commutator, a stepper motor elec-
tronic commutator includes the following three parts:

35

1.   Position sensor
2.   Commutator logic
3.   Electronic coil switches.

5   Comparing this part list with a standard stepper motor controller, we
    see the addition of a position sensor. The rest of the parts are
    equivalent since the commutator logic corresponds roughly in com-
    plexity to the standard sequencing logic. Let us describe the details
    of these three parts.

10

    For a 3-phase uncoupled stepper motor Figure 4 shows how one can
    deduce which of the six-phases (1,...,5,0) the motor is in by using
    three Magnetic Resistor (MR) sensors. The MRs are mounted on the end
    of an E-core and sense the bar's reluctance which changes as the MRs

15 move along the bar's teeth.

    Figure 2 shows an ideal Magnetic Resistor output as a voltage which is
    a sine function of the relative coil-bar position with a period of the
    tooth pitch of the bar (e.g. 2.54 mm). Magnetic resistors (magne-

20 tostrictive resistors) are marketed by Siemens· resistance change as a
    result of magnetic flux and biasing makes the MR a convenient and
    effective proximity transducer for a toothed ferromagnetic bar. The
    three MR sensors are physically mounted to be one third of the bar
    tooth pitch apart. This results in three MR voltages that are 120

25 degrees apart (Figure 4). Using the outputs of three Comparators (C)
    which compare the outputs of each pair of MRs, we obtain three logic
    variables, AB, BC, and CA, which represents a Gray code of the phase
    (half-step) the motor is in.

30 In practice an MR output is a skewed sine function. Nevertheless, the
    exact six motor phases can be obtained by properly adjusting each of
    the MR's gain and offset (4 adjustments) and the MR's skew relative to
    the bar. A second simpler method is to use the zero crossings to
    determine the motor phase (only 3 adjustments). The above adjustments

35 are not very critical, but can be avoided by use of matched MRs. In
    our experience, with MRs that were not well matched, only a single
    initial adjustment was necessary.

It is necessary to do an initial alignment of the MR's sensed motor phase to the actual motor's phase.

In the overall system (Figure 6) the MRs are used for two measure-
5 ments:

o   Motor Phase - coils relative to the bar
o   Motor Position - coils relative to a given point (absolute position).
10

The MR sensor physically measures the magnetic position of the coils relative to the bar. It is an ideal sensor to measure the motor phase, but this sensor can have a noticeable error when interpreting it as the motor "absolute" position. We consider three types of errors:
15

1. Fixed errors which affect the absolute position interpretation of the measurement
2. Variable errors which affect the repeatability of positioning
3. Sensor noise which limits the positioning resolution.
20

The fixed error is because of the accuracy of the bar teeth spacing and in the preferred embodiment is less than 1 mm. This error can be corrected by calibration and use of a lookup table. The variable error is because of the heat expansion of the bar and cannot be calibrated
25 out easily but in the preferred embodiment is limited to 0.1 mm in worst case and typically is less then 0.01 mm. The sensor noise at a band-width of a few kHz limits the position resolution to about 0.0008 mm which gives a dynamic range of 500000 (19 bit) on the 0.5 m bar.
30

When using the MRs to obtain "absolute" motor position, one uses the MR's phase transition counting as the coarse part of the position detection (10 bits in the preferred embodiment) and an A/D (Analog to Digital) conversion of the appropriate MR output to obtain finer
35 resolution (additional 9 bits). The following logic selects the appropriate MR analog output to be read by the A/D converter:

$$MRO = AB \ \overline{BC} + \overline{AB} \ BC$$

$$MR1 = AB \ \overline{CA} + \overline{AB} \ CA \tag{7}$$

5   where the code MR1, MRO is interpreted:

$$01 = MR_A, \quad 10 = MR_B \text{ and } 11 = MR_C.$$

In the preferred embodiment the sampling rate (2 kHz) is not high

10   enough to ensure that the micro-processor did not lose any MR transitions (coarse position). To overcome this problem the MR's phase measurement (AB, BC, CA) is used to update the estimated phase derived from the previously measured motor velocity (see Appendix on velocity estimate) and position. Thus, even with more than two phase changes

15   (half-steps) between samples, the coarse position "count" was not lost.

Commutator Logic

Figure 4 shows the choice of coils to be enabled as a function of six

20   discrete phases (half-steps) and·the desired force direction, Forward or Reverse. This choice follows the commutation principle given before and has quite low force ripple (about 13%) as a function of motor position (Figure 5). The most important property of this commutation function is that the motor will still apply force in the <u>correct</u>

25   <u>direction</u> even with an error of up to plus or minus one incremental phase (one full half-step). This last property is essential for practical operation of a stepper motor commutator due to the inevitable error in the motor phase sensing.

30   The commutation logic for a 3-phase <u>uncoupled</u> stepper motor is represented by the following three logical equations:

$$A = F \ \overline{BC} + \overline{F} \ BC$$

35

$$B = F \ \overline{CA} + \overline{F} \ CA \tag{8}$$

$$C = F \ \overline{AB} + \overline{F} \ AB$$

where A, B, and C are 1 gical variables that represent respectively if coil A or B or C is on, and AB, BC, and CA represent logical values of the comparison between the outputs of the corresponding MRs, i.e. the motor phase in Gray code.

5

For a <u>coupled</u> motor the commutation is somewhat more complex and is given in the Appendix. The added complexity is a result of the coupling between the motor phases which requires the motor to go through twelve half-steps to return to the same orientation of magnetic flux.

10

The sequence of coils enabled follows the standard open-loop sequence of half-steps with an average phase advance in the direction of motion of 1.5 "phases" (between 1 and 2 "phases"). The main difference between our electronic commutation and "standard half-stepping" is

15 that the coil sequencing is generated by closed-loop feedback from the motor phase sensor (MRs in the preferred embodiment) and thus never loses synchrony. In fact, the commutator logic functions as a very fast – at 1.5 m/sec, 3 600 coil switching/sec at the right time are necessary – closed-loop feedback system around the stepper motor. This

20 is the key to the simultaneous achievement of high speed and insensitivity to varying motor load which is almost impossible to achieve with standard sequencing approaches.

The following logic variable S is true if only one coil is energized

25

$$S = F \ (\overline{AB} \ \overline{BC} + \overline{BC} \ \overline{CA} + \overline{CA} \ \overline{AB}) +$$

$$\overline{F} \ (AB \ BC + BC \ CA + CA \ AB) \tag{9}$$

30 In Figure 5 we assumed that the above logic was used to modulate the power amplifier gain (Figure 6) so that the actual force variation (up to 40%) in energizing one or two coils was compensated for using the logic in (9).

35 Phase Advance

In Figure 3 the coil switching circuits are represented by "symb lic" transistors A, B, C. In this paper it suffices to say that these

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circuits can be quite complex depending on the maximum voltage used
and switching times required. Usually "a half-bridge circuit" is a
good choice for an uncoupled V.R. step motor.

5 The practical limitation resulting from the switching circuits is a
limit on the minimum time it takes the current through a coil to be
switched on or off. This limit is because of the coil inductance, L,
and the maximum voltage, V_{max}, the switching circuit can handle
through the equation $\left|di/dt\right|_{max} = V_{max}/L$.

10

Because of this effect, the correct enabling of the coils by the
commutator is effectively delayed as the motor speed increases. When
the motor moves very fast compared to the required switching time of
the coils' current, the commutator feeds the motor with a triangular
15 waveform rather than a square waveform. At this point and beyond, the
effective delay in switching the coils' current is maximum and equal
to 1.5 phases (half-steps).

One can fix this delay problem by appropriately advancing the measured
20 motor phase by using an estimate of the motor velocity. But beyond the
above maximum delay point, more than 1.5 phases advance is not helpful
and the motor force drops monotonically as speed increases (the
maximum current reached in the coils drops).

25 The following is the required logic to achieve phase advance where V_s
is the sign (direction) of the required phase advance and V_1, V_0 is
the binary magnitude in units of half-steps:

$$\widetilde{AB} = V_s\, V_0\, \overline{V}_1\, \overline{BC} + V_s\, \overline{V}_0\, V_1\, CA +$$

30

$$V_0\, V_1\, \overline{AB} + \overline{V}_s\, V_0\, \overline{V}_1\, \overline{CA} +$$

35

$$\overline{V}_s\, \overline{V}_0\, V_1\, BC + \overline{V}_0\, \overline{V}_1\, AB$$

$$\tilde{BC} = V_s\ V_0\ \overline{V}_1\ \overline{CA} + V_s\ \overline{V}_0\ V_1\ AB +$$

$$V_0\ V_1\ \overline{BC} + \overline{V}_s\ V_0\ \overline{V}_1\ \overline{AB} +$$

$$\overline{V}_s\ \overline{V}_0\ V_1\ CA + \overline{V}_0\ \overline{V}_1\ BC$$

$$\tilde{CA} = V_s\ V_0\ \overline{V}_1\ \overline{AB} + V_s\ \overline{V}_0\ V_1\ BC +$$

$$V_0\ V_1\ \overline{CA} + \overline{V}_s\ V_0\ \overline{V}_1\ \overline{BC} +$$

$$\overline{V}_s\ \overline{V}_0\ V_1\ AB + \overline{V}_0\ \overline{V}_1\ CA \tag{10}$$

The function of the phase advance logic is just to do addition of the signed (V_s) magnitude (V_1, V_0) input and the measured phase $(AB, BC, CA$ in Gray code) to obtain the required "advanced-phase" $\tilde{AB}, \tilde{BC}, \tilde{CA}$, also in Gray code. In fact, a standard binary adder could be used (with appropriate wraparound logic) for phase advance if the Gray code were converted to binary (3). This observation is especially important when using more than six phases where the Gray code adder becomes much more complex.

In fact, in the preferred embodiment we could get somewhat better performance if we split every phase into two (use quarter steps rather than half-steps) so we could achieve less ripple in the force as the motor speed increases and attain the ideal maximum phase advance of 3 quarter steps (1.5 half-steps).

To reduce the chip count a Programmed Logic Array (PLA) is used to implement the required logic. For the uncoupled motor one 83S153 PLA was sufficient to contain the commutation logic (8), the one-two coil logic (9) and the phase advance logic (10). For the coupled motor the same PLA was used for phase advance and one-two coil logic, and a second PLA was required for commutation because of the large number of

input/output connections needed and the additional complexity (see Appendix).

Note that one can use the phase advance logic with a simple algorithm
5 to adjust electronically the measured phase of the motor to the actual
motor phase, i.e. the relative relationship between the coil and the
bar.

10 ## Control of a Commutated Step Motor
Digital Proportional-Integral-Derivative Control (PID)
A digital controller for an electronically commutated stepper motor
need only define how much voltage (or current) is to be applied to the
motor at an instant and the direction in which the force is to be
15 generated by the motor. In an ordinary D.C. servo, the same signals
are provided by the magnitude and sign of the driving voltage to the
D.C. motor.

Two linear variable reluctance (V.R.) motors, a coupled phase and an
20 uncoupled phase, have been driven with a PID control law, implemented
on a Motorola 68000 micro-processor and associated D/A's and A/D's
(Figure 6). While this represents the simplest servo strategy, its
implementation for driving a stepping motor is rather novel, and its
good performance will become apparent. Also, the control method is not
25 limited to either linear stepping motors or V.R. stepping motors.

A Motorola 68000 and associated digital and analog ports found on an
RRA card (7) was a convenient host for experiments, although numerous
commercially available micro-processor systems would do. The processor
30 is relegated the task of interpreting motor position, estimating motor
velocity, and synthesizing the control signal. Motor position can be
derived from the MR signals or from a distinct encoder. Velocity could
also be measured from a tachometer, but the added expense and com-
plexity, especially in the linear case, was not justified in the
35 preferred embodiment. The micro-processor can provide an adequate
estimate of velocity.

A Simple PD Controller

The first step in computing the control term is to form the difference between the desired position, X_d, and the actual position of the motor at time k, X_k. Given the velocity of the motor at time k as v_k, a
5   discrete PD control law is then

$$U_k = - K_e (X_k - X_d) - K_v v_k \tag{11}$$

where K, and K_v are gains on positional error and velocity, respec-
10 tively, and U_k is the control.

As no tachometer is present, an estimate, $\hat{v}_k$, must be used for v_k. An easy estimate to form is found by taking a numerical derivative on past position measurements to form an estimate proportional to speed.
15 The s is the first backward difference on position,

$$\hat{v}_k = X_k - X_{k-1} \tag{12}$$

As this is known to be a rather "noisy" estimator, other simple
20 schemes are employed. For example, by fitting 2'nd or 3'rd order polynomials to the data X_k, X_{k-1}, X_{k-2}, ... and then evaluating the derivative of the polynomial at time k, the error in such an estimate can be proven smaller for sufficiently high sample rates. These estimates did in fact improve the performance of the PID servo, run at
25 various sample rates between 0.5 and 2 kHz.

With $U_k = - K_e (X_k - X_d) - K_v \hat{v}_k$, the sign of U_k determines the direction the motor is to accelerate and the magnitude determines the magnitude of voltage. Hence the sign of U_k is output to the commutator
30 for the signal F, the magnitude of U_k is output to the power amplifier to drive the motor, and the value of $\hat{v}_k$ is quantized and provided to the commutator for control of the phase advance (Figure 6).

Compensation for Non-Linearity

35 V.R. motors, unlike D.C. servo motors, have a severe non-linearity, especially at low current levels, below saturation. This n n-linearity is that the static force varies directly with the square of the coil current. As we are trying t "servo" the m tor to a point with a

linear control law, we might expect this non-linearity to adversely affect the performance. The obvious compensation for the non-linearity is not to feed the motor U_k but rather

5
$$u_k = \text{sign}(U_K)\sqrt{|U_K|}$$
(13)

This compensation is correct when current is used as the control variable.

10 In fact, far higher stiffness – larger values of K_e – with good transient response could be achieved with the square root than without it (see Appendix for computation of the square root).

Integral Term and a PID Controller

15 Lastly, to accommodate a zero steady state error criterion, a linear control designer adds an integral term – a sum in the discrete case – to the control, forming the PID control:

$$U_k = - K_e(X_k - X_d) - K_v \hat{v}_k$$

20

$$- K_i \sum_{i=n}^{k} (X_i - X_d)$$
(14)

25 where $n < k$ is typically chosen to be some finite value possibly dependent on $X_k - X_d$, and $\hat{v}_k$. If this control law is asymptotically stable, the value of position X_k as k tends to infinity is X_d exactly.

Unfortunately, an integral controller can suffer from long settling
30 times because of "wind up" of the integral. One scenario is that the integral builds up to overcome a positional error caused by friction. The integral term can not decrease until the sign of the position has changed. Hence, the integral can induce large overshoots.

35 To combat this phenomenon the following anti-windup term for small moves is included, which achieves a much faster settling time.

$$U_k = -K_e(X_k - X_d) - K_v \hat{v}_k k - K_i \sum_{i=n}^{k} (X_i - X_d) - K_w (X_k - X_n) \quad (15)$$

5 The effect of the term $K_w (X_k - X_n)$ is to reduce the effect of the integral term, once the motor starts moving.

The control law used for the V.R. motor then is (13) where U_k is computed by (15).

10

Variable Gains

One can meet difficulties in positioning precisely to 0.01 mm or less on a bar 0.5 m long. Even if the dynamics of the motor were perfectly linear, some other non-linearities are present. One is that the power
15 amplifier can only deliver a limited amount of voltage. To guarantee fast response and to overcome friction to attain high positional accuracies requires large feedback gains, K_e and K_v. Such values are fine for small moves, but when a large move is required, say 0.3 m, these large values of gain immediately saturate the power drive at its
20 maximum output. This effectively drives the motor open-loop for much of the move, and can induce large overshoots. Secondly, as a robot programming language like AML (9) produces sequences of set points (sequences of X_d's) for the actuator to follow, the requirement that the actuator (in this case the linear motor) attain each set point is
25 not really the desired performance. A sequence of evenly spaced points is really meant to drive the actuator at a constant speed, not to jerk it from point to point as quickly as possible. Thirdly, high gains reduce the margin of stability of the controller. Large moves with large gains produce large forces which can induce structural vibra-
30 tions and result in instability.

All three of these problems can be addressed simply by varying the gains K_e and $K_{\hat{v}}$. When the value of $|X_k - X_d|$ or when the value of $\hat{v}_k$ is large, the value of the gains is reduced. This simple strategy
35 alleviates the difficulties of high gains and assures high positional accuracy and fast response. This type of contr 1, so easily imple- mented digitally, is quite cumbersome to effect in an analog con- troller.

Performance

The startling good performance of these rather naive control schemes coupled with the commutator is what makes this entire approach so attractive. For large moves, a PD controller has driven the uncoupled

5 phase motor at speeds over 1.4 m/sec (55 inch/sec) on a 0.5 m (18 inch) bar, and yet stopped it in a critically damped manner.

More important is the positional accuracy and stiffness than can be achieved with this control method. Figure 7 shows the result of

10 commanding the coupled phase linear motor to move in 0.025 mm steps (0.001 inch). The apparent stiffness of this PD controller - the amount of force required to displace the motor from various commanded equilibrium positions a few 0.01 mm - was measured at 90 Nt/mm (450 lb/inch). This is comparable with micro-stepping methods, al-

15 though not quite as good. The transient response is quite good.

In Figure 8 the result of commanding a 300 mm (6 inch) move under direction of AML at a speed of 0.5 m/sec (20 inch/sec) and accelera-tion/deceleration of 10 m/sec/sec (1 g) is shown. The error between

20 the position goal, X_d, and the actual position at each instant nowhere exceeds 5 mm (0.2 inch).

In another application, a PID controller is servoing the uncoupled phase motor to zero error on a 0.002 mm (0.00008 inch) resolution

25 encoder, with a similar transient response.

Control Programs

Derivative of Polynomial for Velocity Estimate

30 While fitting a polynomial and taking a derivative is not an optimum solution to estimating the velocity, it is expedient and requires no modelling of the motor. The first order polynomial's derivative, the slope of the straight line fitting the data X_k and X_{k-1}, is often good enough, but higher order approximations were found to work better.

35

A formula for evaluating the derivative for the unique quadratic through the data X_k, X_{k-1}, X_{k-2} is:

$$3X_k - 4X_{k-1} + X_{k-2} \qquad (16)$$

The equivalent formula for the cubic interpolate is

5

$$11X_k - 18X_{k-1} + 9X_{k-2} - 2X_{k-3} \qquad (17)$$

Each of these formulae provides a signal roughly proportional to the velocity.

10 ## Square Root Computation

Computing a square root in 68000 assembler is not at all difficult. A good initial guess combined with a few Newton-Raphson iterations completes the process. The procedure is outlined here.

15 1. An Initial Guess

Let n be the index of the highest order non-zero bit in the binary representation of the value of $U \geq 0$. For example, suppose $U = 57$ (decimal). Then $n = 6$. The value of n approximates the logarithm of U, hence $n/2$ is approximately the logarith of $\sqrt{U}$. A

20 reasonable first guess to start the process of finding the square root of U is

$$u^0 = 2^{n/2}, \text{ or } 2^{(n-1)/2} \text{ if n is odd} \qquad (18)$$

25

For our example of 57, the first guess then is 8. In general, this initial approximation is within a factor of $\sqrt{2}$ of the exact square root.

2. Newton-Raphson Iteration

30

A recursive iteration for the square root is

$$u^{i+1} = (U/u^i + u^i + 1)/2 \qquad (19)$$

35 For 32 bit integer numbers, the above guess coupled with 3 iterations is sufficient to compute the square r ot to integer precision (16 bit result).

Commutation Logic for a Coupled Motor

The following is the commutation logic required by a magnetically coupled phase stepper motor. The added complexity compared to an uncoupled phase motor is a result of the magnetic coupling between the motor phases which requires the motor to go through twelve half-steps (two cardinal steps) to return to the same direction of magnetic flux. Thus, the added complexity is:

1. Two logical variables are needed for each coil to designate the magnetic flux direction; e.g. for coil AA' in Figure 1 logical variable A is used to designate if "up" direction flux is on or off· while logical variable A is used to designate if "down" direction flux is on or off.

2. Four states (labeled with the two logical variables: T1, T2) are needed to keep track of the twice repeated six half-steps cycle. Variables ST1 (ST2) are used to force T1 (T2) to logic 1 (Set), while CT1 (CT2) are used to force T1 (T2) to logic 0 (Clear).

$$A = F \ \overline{BC} \ \overline{T2} + \overline{F} \ BC \ T1$$

$$B = F \ \overline{CA} \ T2 + \overline{F} \ CA \ \overline{T1}$$

$$C = F \ \overline{AB} \ \overline{T1} + F \ AB \ \overline{T2}$$

$$\tilde{A} = F \ \overline{BC} \ T2 + \overline{F} \ BC \ \overline{T1}$$

$$\tilde{B} = F \ \overline{CA} \ \overline{T2} + \overline{F} \ CA \ T1$$

$$\tilde{C} = F \ \overline{AB} \ T1 + \overline{F} \ AB \ T2 \qquad (20)$$

The states T1 and T2 are updated according to:

$$ST1 = AB \ \overline{BC} \ \overline{CA} \ T2 + AB \ \overline{BC} \ CA \ \overline{T2}$$

$$ST2 = \overline{AB} \ BC \ CA \ \overline{T1} + \overline{AB} \ BC \ \overline{CA} \ T1$$

$$CT1 = AB \ \overline{BC} \ \overline{CA} \ \overline{T2} + AB \ \overline{BC} \ CA \ T2$$

$$CT2 = \overline{AB} \ BC \ CA \ T1 + \overline{AB} \ BC \ \overline{CA} \ \overline{T1}$$

Advantages of the Electromagnetic Actuator System

Important differences between the electromagnetic actuator system of this invention and prior systems are the use of the 3-phase D.C. sensor, the MRs, which directly produce the commutation when processed by commutation logic. There is no need for demodulation nor is there need for a counting of pulses (and potentially miscounting) to achieve position measurement and correct commutation. The MRs and comparators produce the necessary signals for commutation directly· there is no need of an initial motion for alignment to establish the correct commutation. The correct commutation is established, the instant power is turned on.

Secondly, correcting for the square-law non-linearity coupled with the overlapping energization of coils greatly improves the performance. There seems to be little need for further correcting the remaining torque ripple. Since the decision about which coil to enable is handled by the commutator, much of the burden of controlling a step motor has been removed. The control of a commutated step motor is analogous to an ordinary D.C. servo motor. There are some important advantages to such an approach:

*   The stepper motor is no longer restricted to move in steps or half-steps.

Commutation provides a simpler means of positioning between cardinal steps than does conventional micro-stepping. The motor's p sitional accuracy is determined by position feedback and the quality of the control system's design.

* The closed loop control is simple.

A single input, a direction and a magnitude of a voltage, pro-
vides acceleration in a desired direction, independent of the
motor's position. Thus, even simple control laws (e.g. PID) can
achieve high performance.

* More flexible control objectives such as position, velocity, and
force control can be achieved.

* The motor cannot lose synchrony.

Since the commutator automatically enables the proper coils to
generate force in the desired direction, the problem of missed
timing of steps at high speeds in conventional step motor con-
trols does not appear. The motor always applies force in the
prescribed direction even if perturbed by a sudden change in
load.

* The motor will not come to rest in the wrong equilibrium posi-
tion.

The control law defines the equilibrium position. If displaced
the motor will return to the intended equilibrium. With conven-
tional stepping methods, the motor will come to rest at any of
multiple equilibria.

* The motor will not "run away".

Commutation with phase advance guarantees that the motor will not
race away at high speed if perturbed.

* Power dissipation is lowered by one or two orders of magnitude.

Power is expended only when there is a position error. When the
motor settles near or at its desired position, no or very little
control voltage is applied. In contrast, micro-stepping or

traditional stepping requires large currents flowing all the time.

5
* Overlapping the coil energization by alternating between one and two coils reduces the force ripple as the motor's position changes. The low torque ripple helps simple control strategies achieve high precision.

Improvements

10 Some obvious improvements can be made by programming the computer to compensate for known anomalies in the stepper motor, such as temperature rise characteristics or inherent anomalies which affect one type of stepper motor, analogous to the known square-law characteristics of the variable reluctance stepper motor.

15

* Torque ripple compensation.

The system can sense position and adjust power amplifier gain to eliminate the remaining torque ripple as a function of position. A ROM or the MRs can directly provide the compensation as a function of position.

20

* Better control design.

Modelling the commutated motor may lead to improved control law design, incorporating classical design techniques (lead-lag compensation) and modern techniques (observers), and specially customized control laws for replacement stepper motors or for special jobs.

25

30 * A more continuous advance.

It is simple to build a commutator with an advance that has twice the resolution of the one described here. Such a commutator would achieve improved force at high speed.

35 While the invention has been embodied and described as an actuator to control a linear robot, it is obvious that one skilled in the art may

alter it in form and detail for use in other activities, such as the
use of a variable resistance stepper motor, with its high torque and
low inertia, to take the place of ordinary D.C. servo motors where
these characteristics are at a premium. As V.R. step motors are
5 inherently inexpensive to build, and can clearly be controlled to a
high degree of precision, it is conceivable that they might replace
many permanent magnet D.C. motors. And, of course, any application
where stepping motors are now used is a potential candidate for this
method, if the cost of sensing position is justified.

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Claims

1. An electromagnetic actuator system including a stepper motor
having a plurality of windings (11) and power means (20), connected to
5 the windings, to supply power to the windings, characterized by coil
switching means (13), connected to the windings to complete when
selected the enabling circuit from the power means through the se-
lected enabled coil upon activation; programmable logical means (14),
having a plurality of input nodes, having logic to provide position
10 and anomaly corrected switch selection signals as a composite function
of control direction signal and a pattern of present armature (12)
position signals to provide a related pattern of enabling coil selec-
tion signals to the coil switching means; and position sensing means
(13), monitoring the armature position of the stepper motor and
15 feedback connected to the programmable logic means to provide a
pattern of present armature position signal inputs (21) to the pro-
grammable logical means.

2. An electromagnetic actuator system according to Claim 1, wherein
20 a micro-processor (24) is feedback connected to the position sensing
means (13) to receive position inputs, and is connected to the power
means to provide a single amplitude control signal, and further to the
programmable logic means to provide direction and phase advance
signals.
25

3. An electromagnetic actuator system according to Claim 2, wherein
the position sensing means provides digital coarse armature position
feedback signals (22) and analog fine armature position feedback
signals (23) to the micro-processor, and the micro-processor provides
30 control amplitude, control direction and phase advance signals to the
programmable logical means, whereby additional correction is carried
out according to programmed computation by the micro-processor.

4. An electromagnetic actuator system according to Claim 2, wherein
35 the controll computer programming operates with a linear contr 1 law
to provide control amplitude U_k and also includes programming for
square-law compensation, which provides the formula

$$u_k = \text{sign}(U_k)\sqrt{|U_k|}$$

to said power means.

5
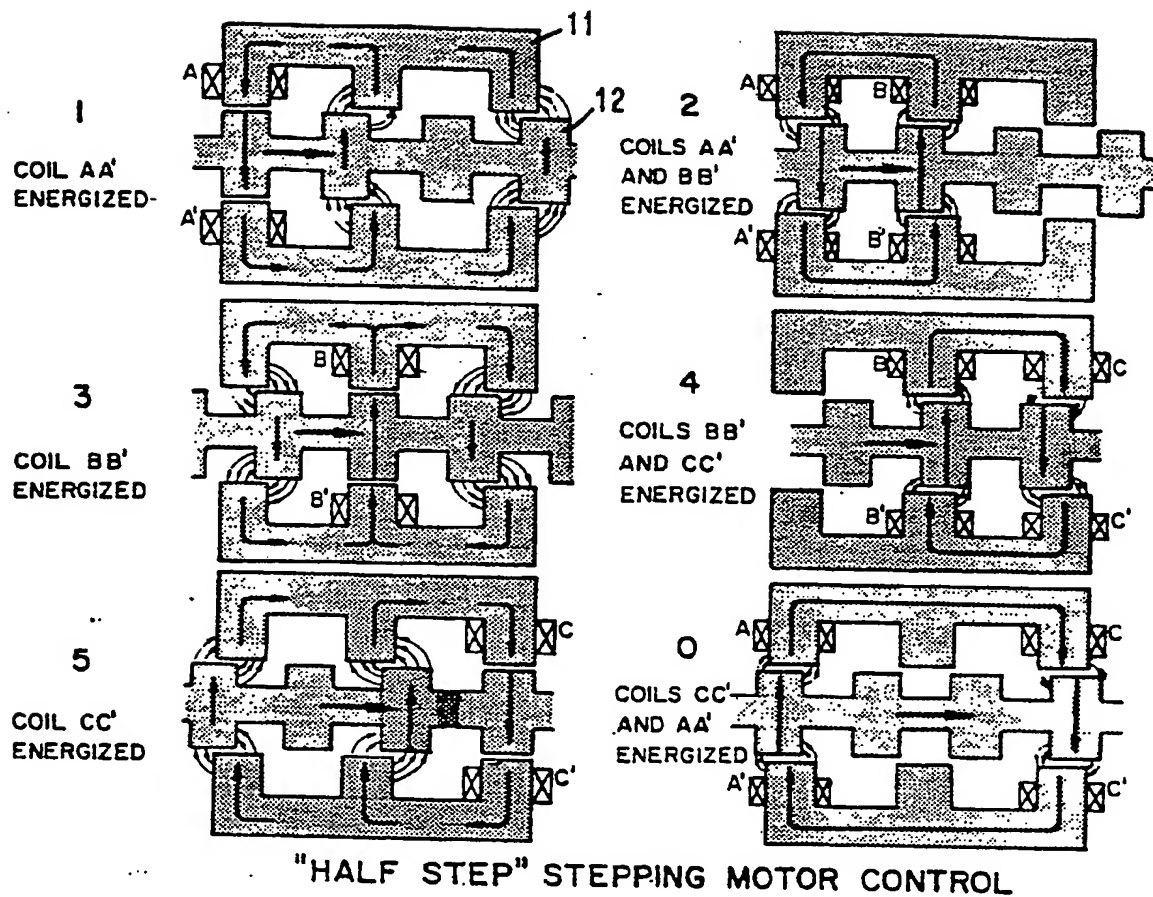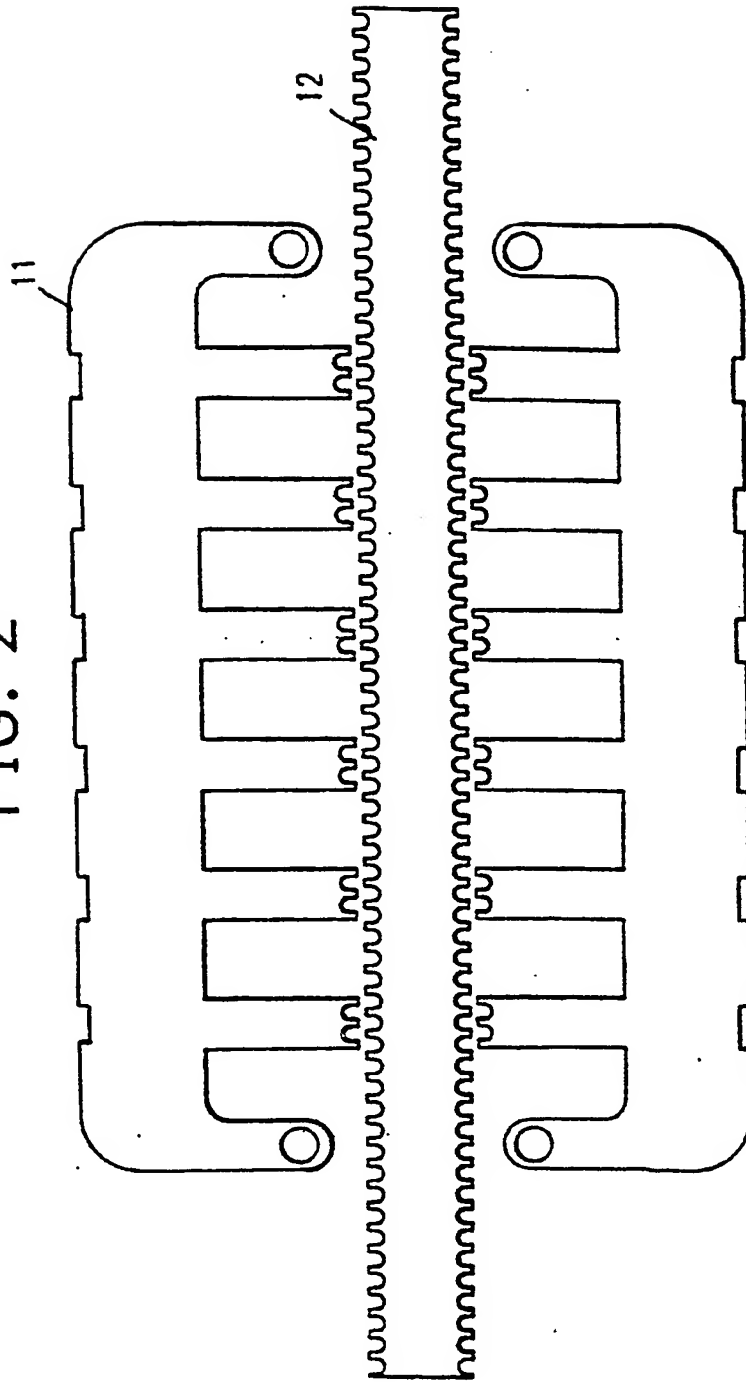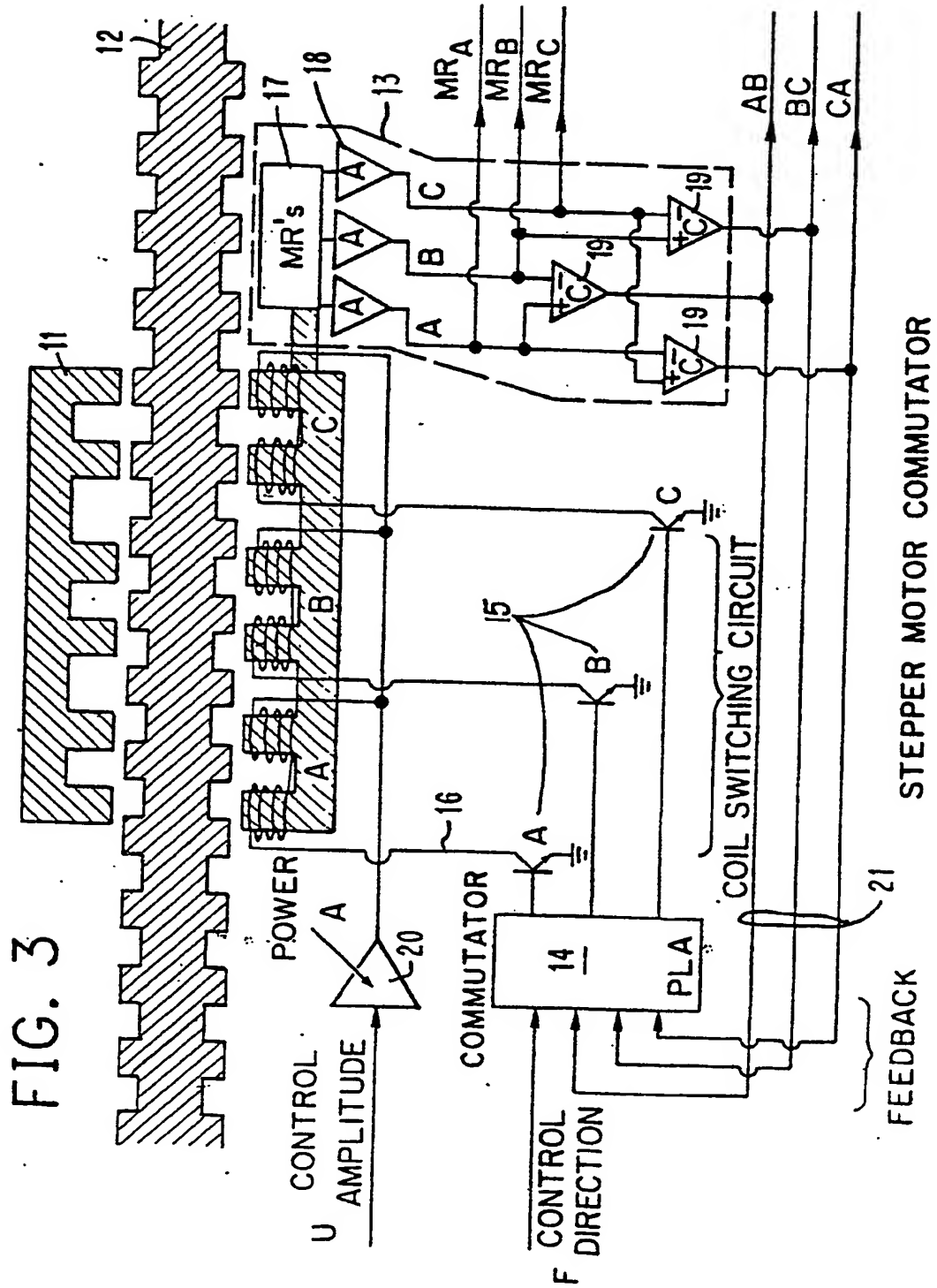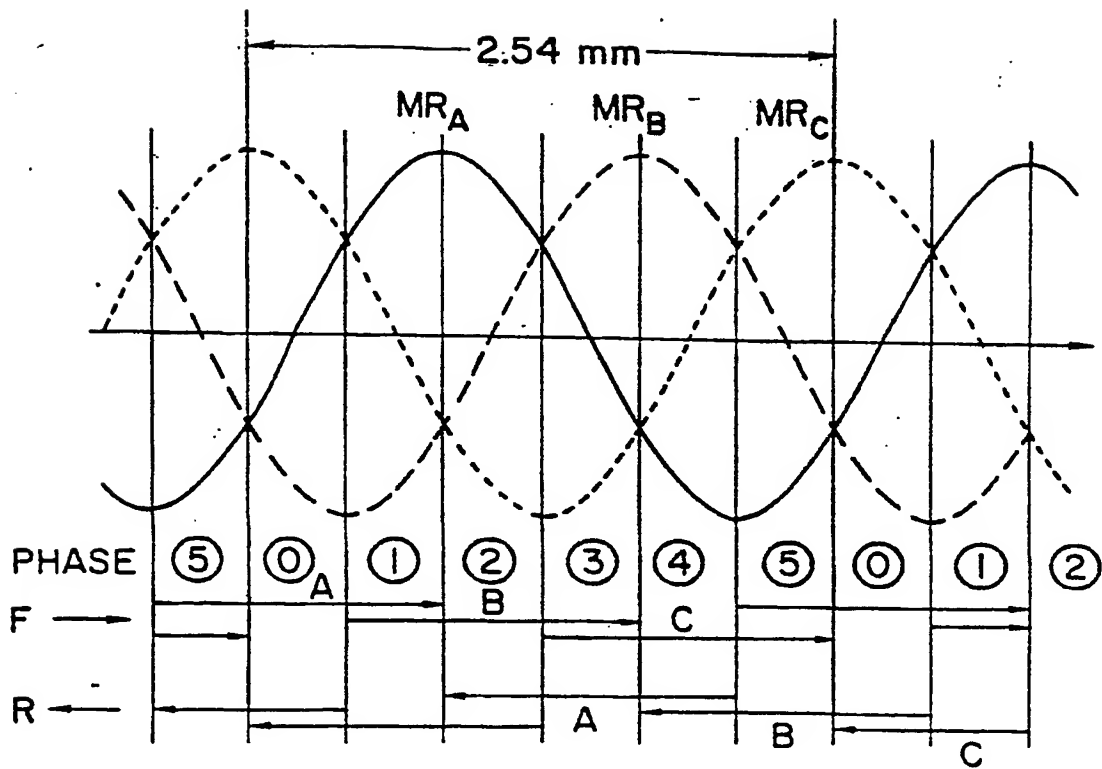
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FIG. 1

1 COIL AA' ENERGIZED

2 COILS AA' AND BB' ENERGIZED

3 COIL BB' ENERGIZED

4 COILS BB' AND CC' ENERGIZED

5 COIL CC' ENERGIZED

O COILS CC' AND AA' ENERGIZED

"HALF STEP" STEPPING MOTOR CONTROL
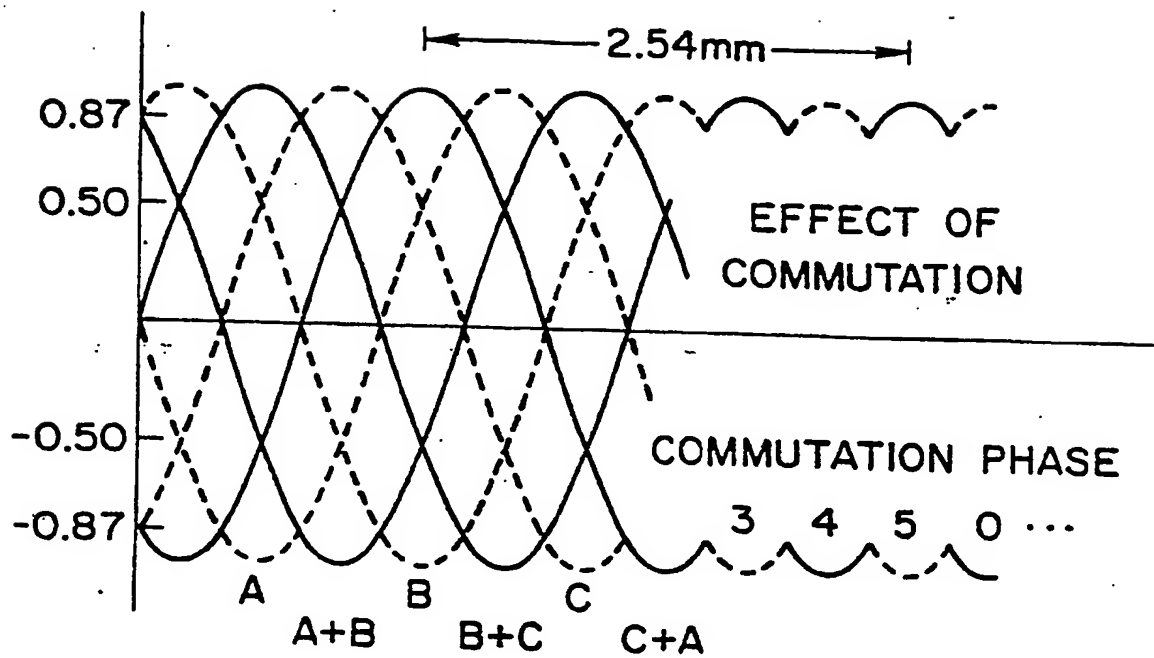
FIG. 2

FIG. 3

STEPPER MOTOR COMMUTATOR

FIG. 4

FIG. 5

STATIC FORCE vs POSITION
FOR FIXED COIL CURRENT

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FIG. 6

STEPPER MOTOR CONTROLLER

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FIG. 7

FIG. 8